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(54) Title: DECODER FOR A NON-COHERENTLY DEMODULATED SIGNAL
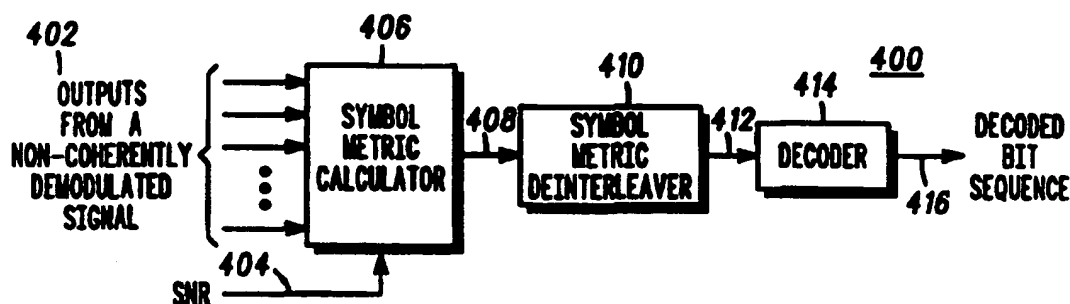
(57) Abstract

Non-coherently demodulated signals are efficiently decoded using a non-complex system that is independent of the fading statistics of the channel. A symbol metric calculator (406) receives a signal-to-noise ratio (404) and symbols from a non-coherently demodulated signal (402) to efficiently produce weighted symbol metrics (408). A symbol metric deinterleaver (410) is then used to deinterleave the weighted symbol metrics. A decoder (414) produces a decoded bit sequence (416) based on the deinterleaved weighted symbol metrics (412).

1

--DECODER FOR A NON-COHERENTLY DEMODULATED SIGNAL--

5   Field of the Invention

The present invention relates to receiver design in a
spread spectrum communication system which employs binary
convolutional coding and orthogonal modulation and, more
10   particularly, to decoding non-coherently demodulated signals.

Background

15   In a spread spectrum communication system, a
modulation technique is utilized in which a transmitted signal
is spread over a wide frequency band within the
communication channel in order to increase the number of
simultaneous users in a communication channel without
20   decreasing performance. The baseband signal (e.g., a voice
signal with a bandwidth of only a few kilohertz) will be
transformed into a signal which occupies and is transmitted
over a frequency band that may be many megahertz wide. This
is accomplished by spreading the signal to be transmitted with
25   a spreading code. These spreading codes include, but are not
limited to, pseudonoise (PN) codes and Walsh codes. A Walsh
code corresponds to a single row or column of the Hadamard
matrix. For example, a dimension 64 channel Hadamard matrix
yields 64 mutually orthogonal Walsh codes. A typical spread
30   spectrum transmission includes a transmitter for expanding
the bandwidth of an information signal and transmitting the
expanded signal and a receiver for recovering the desired
information signal by remapping the received spread spectrum
into the original information signal's bandwidth.

35

2

The receiver in the spread spectrum communication system will generally include a method to demodulate the transmitted signal and a method to decode the demodulated signal. An optimal decoding metric can be computed but is

5   undesirable because of its complexity. The performance of the previous decoding methods is significantly worse than that of the optimal metric, and previous decoding methods require knowledge of the fading statistics of the channel.

10   Accordingly, there is a need for a method, decoder circuit, and system for decoding a non-coherently demodulated signal, wherein the method, decoder circuit, and system are less complex than the optimal method, yield better performance than previous methods, and are independent of the

15   fading statistics of the channel.

Brief Description of the Drawings

20   FIG. 1 is a prior art communication system that uses binary convolutional coding and orthogonal modulation.

FIG. 2 is a flow diagram for implementing a method for decoding a received signal in accordance with the present

25   invention.

FIG. 3 is a flow diagram, shown with greater detail, of the step of calculating k weighted symbol metrics in accordance with the present invention.

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FIG. 4 is a block diagram of a decoding circuit implemented in accordance with the present invention.

3

FIG. 5 is a graphic illustration of a communication
system that uses a decoding circuit for decoding received
signals in accordance with the present invention.

5   FIG. 6 is a block diagram of one embodiment of a receiver
implemented in accordance with the present invention.

FIG. 7 is a flow diagram, shown with greater detail, of
the step of calculating an estimate of the signal-to-noise
10   ratio in accordance with the present invention.

Detailed Description of the Preferred Embodiment

15   Generally, the present invention provides a method,
decoder circuit, and system for decoding a non-coherently
demodulated signal by using a decoder metric which yields
better performance than the prior art, is less complex than the
optimal method, and is independent of the fading statistics of
20   the channel. The symbol metric calculator   receives a signal-
to-noise ratio and symbols from a non-coherently demodulated
signal to efficiently produce weighted symbol metrics. A
symbol metric deinterleaver is then used to deinterleave the
weighted symbol metrics. A decoder produces a decoded bit
25   sequence based on the deinterleaved weighted symbol metrics.
With such a method, apparatus, and system the received signal
is decoded more efficiently than previous techniques.

The present invention is more fully described with
30   reference to FIGs. 1 - 6. FIG. 1, numeral 100, is a prior art
communication system that uses binary convolutional coding
and orthogonal modulation. In the encoding portion (101) of
the communication system, traffic channel data bits (102) are
input to an encoder (104) at a particular bit rate (e.g., 9.6
35   kbits/second). The input traffic channel data bits can include

4

either voice converted to data by a vocoder, pure data, or a combination of the two types of data. Encoder (104) encodes the input data bits (102) into data symbols at a fixed encoding rate with an encoding algorithm which facilitates subsequent

5   maximum-likelihood decoding of the data symbols into data bits (e.g., convolutional or block coding algorithms). For example, encoder (104) encodes input data bits (102) (received at a rate of 9.6 kbits/second) at a fixed encoding rate of one data bit to three data symbols (i.e., 1/3) such that the encoder

10   (104) outputs data symbols (106) at a 28.8 ksymbols/second rate.

The data symbols (106) are then input into a symbol interleaver (108). The symbol interleaver (108) is a block

15   interleaver which interleaves the input data symbols (106) individually as opposed to a block interleaver which groups the data symbols into sets which are kept together during interleaving. In the symbol interleaver (108), the data symbols are individually input into a matrix so that the matrix

20   is filled in a column by column manner. The data symbols are individually output from locations within the matrix so that the matrix is emptied in a row by row manner. Typically, the matrix is a square matrix having a number of rows equal to the number of columns; however, other matrix forms can be chosen

25   to increase the output interleaving distance between the consecutively input non-interleaved data symbols. The interleaved data symbols (110) are output by the symbol interleaver (108) at the same data symbol rate that they were input (e.g., 28.8 ksymbols/second). The predetermined size of

30   the block of data symbols defined by the matrix is derived from the maximum number of data symbols which can be transmitted at a predetermined chip rate within a predetermined length transmission block. For example, if data symbols (106) are output from the encoder (104) at a 18.8

35   ksymbols/second rate, the chip rate for transmitting those

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data symbols (106) is 228.8 ksymbols/second. If, for example,
the predetermined length of the transmission block is 20
milliseconds, the predetermined size of the block of data
symbols is 28.8 ksymbols/second times 20 milliseconds which
5   equals 576 data symbols. A matrix of dimension 18 by 32
exactly accommodates these 576 data symbols.

The interleaved data symbols (110) are then input to a
mapper (112). The mapper (112) derives a sequence of fixed
10   length orthogonal codes (114) (e.g., 64-ary Walsh codes) from
the interleaved data symbols (110). For example, in 64-ary
orthogonal signaling, the interleaved data symbols (110) are
grouped into sets of six in order to select one of the 64
orthogonal codes to represent the set of six data symbols.
15   These 64 orthogonal codes preferably correspond to Walsh
codes from a 64 by 64 Hadamard matrix wherein a Walsh code
is a single row or column of the matrix. The mapper (112)
outputs a sequence of Walsh codes (114) which correspond to
the input data symbols (110) at a fixed symbol rate (e.g., 307.2
20   ksymbols/second).

The sequence of Walsh codes (114) is output from the
encoding portion (101) of the communication system and input
to a transmitting portion (116) of the communication system.
25   The sequence (114) is prepared for transmission over a
communication channel by a modulator (117). Subsequently,
the modulated sequence is provided to an antenna (118) for
transmission over the communication channel (120).

30   The modulator (117) preferably prepares the sequence
(114) for direct sequence code d:vided spread spectrum
transmission by spreading the sequence (114) with a long
spreading code (e.g., pseudonoise (PN)). The spreading code is a
user specific sequence of symt   or unique user code which is
35   output at a fixed chip rate (e.g. ι.228 Mchips/second). The

6

long spreading code allows the transmission of different users
to be separately demodulated even though the transmissions
occupy the same bandwidth. The unique user code also
enhances the security of the communication in the
5   communication channel by scrambling the encoded traffic
channel Walsh codes. In addition, the code spread Walsh
symbols are spread by a pair of short spreading codes (i.e.,
short when compared to the long spreading code) to generate
separate I-channel and Q-channel code spread sequences. The
10   I-channel and Q-channel code spread sequences are used to bi-
phase modulate a quadrature pair of sinusoids. The modulated
sinusoids are summed, bandpass filtered, translated to an RF
frequency, amplified, filtered and radiated by an antenna (118)
to complete transmission of the traffic channel data bits
15   (102) in a communication channel (120).

A receiving portion (122) of the communication system
receives the transmitted spread spectrum signal from over the
communication channel (120) through antenna (124). The
20   received signal is sampled into data samples by demodulator
(126). Subsequently, the data samples (128) and (129) are
output to the decoding portion (130) of the communication
system.

25   The demodulator (126) preferably samples the received
spread spectrum signal by filtering, demodulating, translating
from the RF frequency, and sampling at a predetermined rate
(e.g., 1.2288 Msamples/second). Subsequently, the in-phase
sampled signal and quadrature sampled signal are
30   independently despread by correlating the received sampled
signals with the short spreading codes and the long spreading
code. As the spreading factor of the long code is four, the
despread samples are partitioned into blocks of four and
summed. Combining samples in this manner reduces the

7

sample rate by a factor of four to a sample rate equal to the
Walsh chip rate which is 307.2 chips/second.

5   The in-phase (128) and quadrature (129) sampled signals
are independently input to a decoding portion (130) of the
communication system which non-coherently detects/decodes
the sampled signals (128). In order to decode the sampled
signals (128 and 129), predetermined length groups (e.g., 64
sample length groups) of sampled signals are independently
10   input to orthogonal code transformers (132 and 134,
respectively) (e.g., fast Hadamard transformers) . The ·
orthogonal code transformers (132 and 134) output a plurality
of output signals (133 and 135, respectively) (i.e., when 64
sample length groups are input, then 64 transformer output
15   signals are generated). Each transformer output signal
corresponds to a measure of confidence that the sampled
signals correspond to a particular orthogonal code from within
a set of mutually orthogonal codes. In addition, each
transformer output signal has an associated index which
20   indicates which particular orthogonal code within the set of
mutually orthogonal codes corresponds to the transformer
output signal (i.e., when 64 sample length groups are input, the
6 bit index is the sequence of six data symbols which map into
the particular length 64 bit length orthogonal code).
25   Subsequently, each transformer output signal within the
groups of transformer output signals (133 and 135) is squared
by the transformer output squaring mechanisms (136 and 138,
respectively). A set of decision values (142), equal in number
to the number of Walsh codewords, is then generated by adding
30   mechanism (140) which adds together each pair of squared
transformer output signals (i.e., one from each of the
transformer output signal squaring mechanisms (136 and 138))
having an associated index which indicates that the
transformer output signals correspond to the same orthogonal
35   code.

8

The group of decision values (142) and associated index data symbols are input to a selection mechanism (144) which selects the maximum decision value from the group of decision
5   values (142). The selected decision value (146) is input to a metric computation mechanism (150) which scales the selected decision value to a value (154) which is used to form metrics for the individual data symbols which are subsequently used to form transition metrics for maximum-
10   likelihood decoding techniques. The index (a sequence of binary symbols) associated with the selected decision value (148) is determined by the mapping mechanism (152). Multiplier (158) multiplies each index symbol (156) by the scaling factor (154) to form a symbol metric (160) for each
15   data symbol. The symbol metrics are formed at a rate equal to the data symbol rate at the output of the encoder.

The data symbol metrics (160) are then input into a symbol metric deinterleaver (162) which deinterleaves the
20   data symbol metrics (160) at the individual symbol level. In the symbol metric deinterleaver (162), the symbol metrics are individually input into a matrix which defines a predetermined size block of soft decision data. The matrix of symbol metrics is filled in a row by row manner. The deinterleaved symbol
25   metrics (164) are individually output from locations within the matrix so that the matrix is emptied in a column by column manner. The deinterleaved symbol metrics (164) are output by the symbol metric deinterleaver (162) at the same rate that they were input (e.g., 28.8 kmetrics/second). The dimension of
30   the symbol metric deinterleaver is the same as the dimension of the symbol interleaver since it inverts the operation of the symbol interleaver and there is one symbol metric for each data symbol.

9

The deinterleaved symbol metrics (164) are input to a decoder (166) which uses maximum-likelihood decoding techniques to estimate the traffic channel data bits (168). The maximum-likelihood decoding techniques may be

5   augmented by using an algorithm which is substantially similar to a Viterbi decoding algorithm. The decoder (166) uses a group of the data symbol metrics (154) to form a metric for each possible state transition within the maximum-likelihood sequence estimation decoder (166). The

10   number of deinterleaved symbol metrics (164) used to form each set of transition metrics is equal to the number of data symbols (106) generated at the output of the convolutional encoder (104) from each input data bit (102). The number of transition metrics in each set is equal to two raised to the

15   power of the number of data symbols (106) generated at the output of the convolutional encoder (104) from each input data bit (102). For example, when a rate 1/3 convolutional encoder is used in the transmitter, three data symbols (106) are generated from each input data bit (102). Thus decoder (166)

20   uses groups of three deinterleaved symbol metrics (164) to form eight transition metrics for use at each time state in the maximum-likelihood sequence estimation decoder (166). The estimated data bits (168) are generated at a rate equal to the data rate at the input to the decoder.

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Thus a prior art communication system for using binary convolutional coding and orthogonal modulation in encoding and decoding has been described above with reference to FIG 1, numeral 100. In summary, the prior art communication system

30   includes a first portion (101 and 116) which encodes input data bits into data symbols, interleaves the data symbols in a symbol by symbol manner, maps the interleaved symbols into orthogonal codes, modulates and transmits the orthogonal codes over a communication channel. The prior art

35   communication system further includes a second portion (122

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and 130) which receives and demodulates a signal from over
the communication channel, transforms groups of samples of
the demodulated signal into a group of measures of confidence
that each particular group of samples corresponds to a

5   particular orthogonal code from within a set of mutually
orthogonal codes, selecting the one largest measure of
confidence and an index data symbol which identifies the
particular orthogonal code corresponding to the selected
measure of confidence, generating data symbol metrics from

10   each selected measure of confidence and associated index data
symbol, deinterleaving the data symbol metrics within each
received transmission block, subsequently generating
transition metrics from groups of deinterleaved individual
symbol metric, and generating estimated data bits from the

15   transition metrics, by using maximum-likelihood decoding
techniques.

FIG. 2, numeral 200, is a flow diagram for implementing
a method for decoding a received signal in accordance with the

20   present invention. If a signal-to-ratio (SNR) is not available,
an SNR estimate is calculated based on M outputs from a non-
coherent demodulation of a received signal (208). Then, k
weighted symbol metrics (wherein k is a number of data
symbols) are calculated, for each M-ary symbol, using the

25   signal-to-noise ratio or the SNR estimate of the received
signal and M outputs from the non-coherent demodulation of
the received signal (202). Then, the weighted symbol metrics
are deinterleaved (204) and input into the decoder. The
decoder uses the deinterleaved symbol metrics to produce a

30   decoded bit sequence(206). More description of these steps is
given below. Where the transmitted signal is convolutionally
encoded, the decoder is typically a Viterbi decoder. For a
block encoded signal, corresponding soft-decision decoding
block decoding algorithms can be employed. The output of the

decoder is the binary input sequence corresponding to the
decoded codeword.

FIG. 3, numeral 300, is a flow diagram of the step of
5   calculating k weighted symbol metrics, shown with greater
detail, in accordance with the present invention. First, for
each of the k data symbols corresponding to particular
received M-ary codeword, a maximum one output is selected
that is the largest of the M outputs representing symbols for
10   which the corresponding data symbol is a one (302), and a
maximum zero output is selected that is the largest of the M
outputs representing symbols for which the corresponding data
symbol is a zero (304). Then for each of the k data symbols, a
difference is taken between the maximum zero output and the
15   maximum one output (306). The k differences corresponding to
the k data symbols are subsequently multiplied by a function
of the signal-to-noise ratio (SNR) for the power control group
to produce k weighted symbol metrics (308). For many
applications the SNR may need to be estimated. The
20   multiplication of the difference by the estimate of the SNR is
an approximation of the log-likelihood ratio. Multiplication by
a function of the estimate of the SNR, such as the square root,
may also be used as an approximation of the log-likelihood
ratio to yield a better decoder performance. More description
25   of these steps is given below.

FIG. 4, numeral 400, is a block diagram of a decoding
circuit implemented in accordance with the present invention.
The decoding circuit (400) includes a symbol metric calculator
30   (406), symbol metric deinterleaver (410), and a decoder (414).
The symbol metric calculator (406) receives outputs from a
non-coherently demodulated signal (402) and a signal-to-noise
ratio (404). The outputs from a non-coherently demodulated
signal (402) will typically either be the squared outputs of a
35   single M-ary noncoherent demodulator, or the square-law

12

combination of the outputs of several M-ary demodulators.
This latter case, in which the outputs of several M-ary
demodulators are combined would correspond to instances
when diversity is available to the receiver, such as in the case
5   of the IS-95 system where the multipath nature of the channel
provides diversity. The signal-to-noise ratio (404) is required
in the calculation of the weighted symbol metric (408). The
signal-to-noise ratio weighting employed by the weighted
symbol metric (408) is only necessary for a channel in which
10   the signal-to-noise ratio is time-varying. The signal-to-noise
ratio is approximately constant over a power control group
(six Walsh symbols), allowing estimation of the average
signal-to-noise ratio for the power control group. In systems
in which the signal-to-noise ratio varies more slowly or more
15   rapidly, the method for estimating the signal-to-noise ratio is
modified accordingly. The symbol metric calculator (406)
produces weighted symbol metrics (408) based on the outputs
from a non-coherently demodulated signal (402) and the SNR
(404). The weighted symbol metrics provided by the symbol
20   metric calculator are passed to the symbol metric
deinterleaver (410) to produce deinterleaved weighted symbol
metrics (412). The deinterleaved weighted symbol metrics
(412) are passed to the decoder (414). The decoder (414) uses
the deinterleaved weighted symbol metrics to produce a
25   decoded bit sequence(416). Where the transmitted signal is
convolutionally encoded, as in IS-95, the decoder (414)
typically is a Viterbi (maximum-likelihood) decoder. If some
other type of block encoding is performed by the transmitter,
corresponding block decoding methods are utilized. These
30   block decoding methods include both maximum-likelihood
decoding and methods which provide near maximum-likelihood
decoding performance. The output (416) of the decoder (414)
is a binary sequence corresponding to the decoded codeword.

13

FIG. 5, numeral 500, is a graphic illustration of a communication system that uses a decoding circuit (400) for decoding received signals in accordance with the present invention. Examples of communication systems are cellular

5   telephone systems and personal communication systems. The communication system (500) includes communication units (502), a base station (508), and interfaces to the public telephone network (514). The base station (508) uses a forward link (506) from an antenna (516) as the channel

10 resource to communicate with the communication unit (502). The communication unit (502) uses a reverse link (504) to the antenna (516) as the channel resource to communicate with the base station (508). The base station (508) is shown with a receiver (510) that utilizes a decoding circuit (400) in

15 accordance with the present invention for receiving communications from the communication units (502) and a transmitter (512) for sending communications to the communication units (502).

20   FIG. 6, numeral 600, is a block diagram of one embodiment of a receiver implemented in accordance with the present invention. The system (600) includes one or more M-ary demodulators (602) for demodulating the received signal (601). Each M-ary demodulator (602) non-coherently

25 demodulates a separate multipath component of the received signal (601) and produces M outputs (603). The M outputs (603) are combined using a square-law combiner (604). The output of the square-law combiner (604) is then delayed by delay element (608) before being input to the symbol metric

30 calculator (406). The symbol metric calculator (406) receives the delayed output (402) from the square-law combiner (604) and a signal-to-noise ratio estimate (SNR) (404) from the SNR calculator (606). The signal-to-noise ratio estimate (404) is derived from the output of the square-law combiner (604).

35 The symbol metric calculator (406) uses the SNR estimate

14

(404) and the delayed output (402) from the square-law
combiner (604) to produce the weighted symbol metrics (408).
The weighted symbol metrics (408) are deinterleaved by the
symbol metric deinterleaver (410). The output of the symbol
5   metric deinterleaver (412) is used by the decoder (414) to
produce the decoded bit sequence(416).

The received signal (601) is representative of the
signals received by the antenna (516) in FIG. 5, numeral 500.
10   The communication channel is generally a multipath channel so
that the received signal is the linear sum of several delays of
the transmitted signal. In addition to introducing multipath,
the channel also corrupts the signal by adding additive white
Gaussian noise of one-sided spectral density N_0. As a
15   consequence of the long code pseudonoise (PN) spreading
employed by the transmitter, multipath components having
delays separated by delays caused by more than one chip may
be demodulated without interfering with each other.

20   FIG. 7, numeral 700, is a flow diagram of the step of
calculating an estimate of the signal-to-noise ratio, shown
with greater detail, in accordance with the present invention.
First, an estimate of signal plus noise is defined as a
maximum of the M outputs (702). Then, an estimate of the
25   noise is defined as the average of all of the M outputs except
for the maximum of the M outputs (704). Then, an estimate of
the signal is calculated as the estimate of signal plus noise
minus the estimate of the noise (706). Then, an estimate of
the signal-to-noise ratio is calculated by dividing the
30   estimate of the signal by the estimate of the noise (708). An
optional final step, of calculating an approximately unbiased
signal-to-noise ratio estimate by subtracting, from the
estimate of the signal-to-noise ratio, a bias value consistent
with the estimate of the signal-to-noise ratio (710), may also
35   be added. More description of these steps is given below.

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There are several unique aspects of this system with respect to the symbol metric calculator (406). First, the power control is employed on the reverse link in order to
5   maintain the power received by the base station at an approximately constant level. Power control is implemented by partitioning the reverse link transmission into power control groups. Typically, a power control group consists of 6 Walsh symbols so that there are 16 power control groups per
10   20 millisecond traffic frame. The base station measures the power received over a power control group and then commands the mobile to increment or decrement the transmitted power by a fixed increment (e.g., 1 dB), depending on whether the measured power was above or below the desired threshold.
15   Note that the base station must estimate the signal-to-noise ratio for the power control group in order to implement power control.

In a communication system in which the SNR is not
20   constant over the duration of a frame, the decoded performance, i.e., frame error rate and bit error rate can be greatly enhanced if the instantaneous SNR is known and is exploited in the decoding process. In the present invention, the SNR is approximately constant over the duration of the
25   power control group is utilized. The SNR is only approximately constant for two reasons. The first reason is that while the transmitter power is held constant for the duration of the power control group, the fading due to the channel is a continuous random process. Thus, although the transmitted
30   power of the desired transmission is held constant for the duration of the power control group, the energy received by the base station may not be constant. The second reason that the SNR is only approximately constant over the duration of a power control group is that the interference is also time-
35   varying. If the power control groups of the interfering

16

transmissions are aligned with that of the desired power control group, the transmitted energy of the interference is constant over the duration of the power control group.

However, the interference is also subject to a time varying
5   fading process so that the energy of the interference received at the base is not constant. However, while the SNR may vary within the power control group, there is much less variation of the SNR within the power control group than within the 16 power control groups which constitute a frame.

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Let the vector $z^{(p)} = (z_1^{(p)}, z_2^{(p)}, K \ z_M^{(p)})$ denote the vector output of the square-law combiner (604) for the p-th M-ary symbol within the power control group, $1 \le p \le 6$. We then have the estimates

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$$\hat{N}_0^{(p)} = \sum_{\{k:z_k^{(p)} \ne z_{max}^{(p)}\}} z_k^{(p)} / 63 \qquad \hat{E}_W^{(p)} = z_{max}^{(p)} - \hat{N}_0^{(p)},$$

of the noise and signal, respectively, for the p-th M-ary symbol, where $z_{max}^{(p)} = \max\{z_k^{(p)}\}$. One estimate $\hat{E}_W / \hat{N}_0$ for the
20   power control group is then given by

$$\hat{E}_W / \hat{N}_0 = \left(\sum_{1 \le p \le 6} \hat{E}_W^{(p)} \right) \Big/ \left(\sum_{1 \le p \le 6} \hat{N}_0^{(p)} \right).$$

The rationale for this estimate is that the largest output
25   of the square-law combiner (604) usually corresponds to the transmitted Walsh symbol. Of course, the largest output does not always correspond to the transmitted Walsh symbol, especially when the SNR is small. For this reason, the mean of the estimate $\hat{E}_W$ is always greater than the actual value E_W,
30   and the mean of the estimate $\hat{N}_0$ is always less than the actual value N_0. The bias of $\hat{N}_0$ is very small due to the number of

17

samples used to form the estimate. For this same reason, the variance of $\hat{N}_0$ is small also. Since the bias of the estimate of $\hat{E}_W$ is a function of N_0, an approximately unbiased estimate $\hat{E}_{W,u}$ can be achieved by subtracting the bias consistent with

5   the estimated noise level $\hat{N}_0$ from the estimate of $\hat{E}_W$. The estimates $\hat{E}_{W,u}$, $\hat{N}_0$, and SNR estimate $\hat{E}_{W,u} / \hat{N}_0$ constitute the contents of the SNR calculator (606).

There is an alternative method for estimating the SNR

10   which yields better performance in some instances. In this alternative method, the noise is estimated as before; that is, all the outputs of the demodulator other than the largest are based upon noise only. An estimate of the noise density is computed by averaging the demodulator outputs for the entire

15   power control group which are in correspondence with noise. Note that since the number of noise outputs in a power control group is typically 378, the noise estimate is corrupted only slightly if the largest demodulator output for one or more symbols of a power control group do not correspond to the

20   transmitted signal. The noise estimate is used to normalize the largest demodulator output. The maximum-likelihood estimate of the signal-to-noise ratio is then given by $\hat{E}_{W,ml} / \hat{N}_0$,

25   where

$$\hat{E}_{W,ml} / \hat{N}_0 = \max_x \; \text{prob}(z_{max} | N_0 = \hat{N}_0, E_W = x).$$

The conditional density of the largest demodulator output as a

30   function of the normalized signal energy depends only on the number of multipath components combined in the receiver. The receiver has knowledge of the number of multipath components combined because the receiver controls this value. The

18

maximum-likelihood estimate of the signal-to-noise ratio can be implemented using a table lookup procedure. A separate table is required for each multipath component.

5   With respect to the symbol metric calculator (406), typically for an arbitrary M-ary symbol q with binary representation $(b_{q,1}, b_{q,2}, ..., b_{q,6})$ a metric is assigned to each bit of the binary representation of q for use in the Viterbi decoder (414).

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The optimal decoding metric for a Viterbi decoder is the log-likelihood ratio, defined below. For each index i, $1 \leq i \leq 6$, let S_0^i denote the set of M-ary symbols for which the binary representation has a 0 in the i-th location and let S_1^i denote

15   the complementary set. The log-likelihood ratio for a binary symbol b_i in the i-th location of the binary representation of an M-ary symbol is given by

$$
\log\left(\frac{p_0^i(b_i, z)}{p_1^i(b_i, z)}\right) = (-1)^{b_i} \log\left(\frac{\sum_{j \in S_0^i} z_j^{(m-1)/2} I_{m-1}(2\sqrt{2z_j E_w / N_0})}{\sum_{k \in S_1^i} z_k^{(m-1)/2} I_{m-1}(2\sqrt{2z_k E_w / N_0})}\right).
$$

where $I_a(x)$ denotes the a-th order modified Bessel function of

20   the first kind, and z is the output of the square-law combiner (604).

This optimal receiver metric is too complicated to implement due to the number and complexity of the

25   calculations involved. An alternative metric which is simpler to implement is set forth below. Let $z_{j,max}^i$ denote the largest element of the set S_j^i. Refer to the log-likelihood ratio above. Since the summations of the numerator and denominator are dominated by the terms corresponding to the largest elements

30   of S_0^i and S_1^i, respectively, either of two alternatives set forth below for simplifying the metric computation may be used.

19

The first alternative is approximation of the numerator and the denominator by their maximum terms. However, since the resulting approximation will depend only on $z^i_{0,max}$ and $z^i_{1,max}$, a more direct approach is to determine a log-likelihood ratio for the binary symbol b_i which exploits only the quantities $z^i_{0,max}$ and $z^i_{1,max}$. The log-likelihood ratio for b_i based only on $z^i_{0,max}$ and $z^i_{1,max}$ is given by

$$\log\left(\frac{p^i_0(b_i, z^i_{0,max} = x, z^i_{1,max} = y)}{p^i_1(b_i, z^i_{0,max} = x, z^i_{1,max} = y)}\right) =$$

$$(-1)^{b_i} \log\left(\left(\frac{f^m_{nc}(x)F^m_c(x) + (\frac{M}{2} - 1)F^m_{nc}(x)f^m_c(x)}{f^m_{nc}(y)F^m_c(y) + (\frac{M}{2} - 1)F^m_{nc}(y)f^m_c(y)}\right)\frac{f^m_c(y)F^m_c(y)}{f^m_c(x)F^m_c(x)}\right).$$

In the above,

$$f^m_c(x) = \frac{x^{m-1}e^{-x}}{(m-1)!}, \qquad F^m_c(x) = 1 - e^{-x}\sum_{j=1}^{m}\frac{x^{j-1}}{(j-1)!}$$

and

$$\left(x / 2(E_w / N_0)\right)^{(m-1)/2} e^{-(x + 2E_w/N_0)} I_0\left(2\sqrt{2xE_w / N_0}\right),$$

$$F^m_{nc}(x) = 1 - Q_m\left(2\sqrt{E_w / N_0}, \sqrt{2x}\right),$$

where Q_m denotes the generalized Marcum's Q-function.

The simplified log-likelihood ratio is still difficult to evaluate due to the presence of both exponentials and the generalized Marcum's Q function. However, the following simple expression is a good approximation to the log-likelihood ratio:

$$\log\left(\frac{p^i_0(b_i, z^i_{0,max} = x, z^i_{1,max} = y)}{p^i_1(b_i, z^i_{0,max} = x, z^i_{1,max} = y)}\right) \approx c_1(-1)^{b_i}\left(E_w / N_0\right)(x - y)$$

20

The constant c_1 has no effect on which path is selected by the decoder and can therefore be set arbitrarily in the implementation. Insertion of the estimated SNR into the above expression yields the *weighted symbol metric* for b_i, which is

5   given by

$$\log\left(\frac{p_0^i(b_i, z_{0,\max}^i = x, z_{1,\max}^i = y)}{p_1^i(b_i, z_{0,\max}^i = x, z_{1,\max}^i = y)}\right) \approx c_1 (-1)^{b_i} \left(\hat{E}_{w,u} / \hat{N}_0\right)(x - y).$$

The symbol metric calculator (406) calculates the weighted symbol metric using the inputs from both the delay
10   element (608) and the SNR estimate (404) provided by the SNR calculator (606). The delay element inserts a delay of exactly one power control group (six Walsh symbols) so that the SNR estimate required by the symbol metric calculator (406) may be calculated by the SNR calculator (606). The output of the
15   symbol metric calculator is then passed on to the symbol metric deinterleaver (410). The function of the symbol metric deinterleaver is inversion of the operation performed by the symbol interleaver at the transmitter.

20   After deinterleaving, the metrics are passed to the decoder (414) which will in general be a Viterbi decoder. As the metric is defined here, the Viterbi decoder selects the codeword having the largest metric. That is, the Viterbi decoder decodes that sequence $\{b_i\}_{i=1}^{576}$ belonging to the
25   convolutional code for which the sum of the metrics of the individual binary symbols is largest. The output (416) of the decoder (414) is the binary input sequencing corresponding to the decoded codeword.

30   The present invention provides a method, decoder circuit, and system for decoding a received signal in a communication system. With the present invention, it is possible to significantly reduce, relative to prior art solutions, the

21

signal-to-noise ratio required to achieve a given bit error probability or frame error probability. In addition, this method and apparatus is less susceptible to degradation resulting from operation in environments in which the

5   background noise N_0 is time varying, and a time varying background noise level N_0 is characteristic of the environment in which this invention will be implemented. Additionally, it should be noted that any reduction of signal-to-noise ratio required to achieve a given level of

10 communication reliability corresponds directly to an increase in the capacity of the CDMA system in which this invention operates. Finally, it should be noted that the method is sufficiently non-complex to enable computation of the weighted symbol metric at the receiver.

15

Although exemplary embodiments are described above, it will be obvious to those skilled in the art that many alterations and modifications may be made without departing from the invention. Accordingly, it is intended that all such

20 alterations and modifications be included within the spirit and scope of the invention as defined in the appended claims.

25   We claim:

22

1.   A method for efficiently decoding a non-coherently
demodulated signal in a communication system, wherein the
non-coherently demodulated signal includes a plurality of M-
ary symbols each with M outputs and corresponding to k data
5   symbols, where M equals 2^k, and the non-coherently
demodulated signal is associated with a signal-to-noise ratio,
the method comprising the steps of:

1A)   calculating k weighted symbol metrics, for each M-ary
10   symbol, based on the signal-to-noise ratio and the M outputs
to produce a plurality of weighted symbol metrics;

1B)   deinterleaving, using a symbol metric deinterleaver, the
plurality of weighted symbol metrics to produce a plurality of
15   deinterleaved weighted symbol metrics; and

1C)   decoding, using a decoder, the plurality of deinterleaved
weighted symbol metrics to produce a decoded signal.

2.   The method of claim 1, wherein in step a), the signal-to-noise ratio is an estimate.

3.   The method of claim 1, wherein the method further
5   comprises an initial step of calculating an estimate of the signal-to-noise ratio based on the M outputs, and where selected, one of 3A-3B:
        3A)   wherein calculating an estimate of the signal-to-noise ratio uses the maximum-likelihood estimate (MLE); and
10   3B)   wherein calculating an estimate of the signal-to-noise ratio further comprises the steps of:
        3B1) defining an estimate of signal plus noise as a maximum of the M outputs;
        3B2) defining an estimate of the noise as the
15   average of all of the M outputs except for the maximum of the M outputs;
        3B3) calculating an estimate of the signal as the estimate of signal plus noise minus the estimate of the noise; and
20   3B4) calculating an estimate of the signal-to-noise ratio by dividing the estimate of the signal by the estimate of the noise;
and, where selected,
        3B5)   wherein the method further comprises a final
25   step of calculating an approximately unbiased signal-to-noise ratio estimate by subtracting, from the estimate of the signal-to-noise ratio, a bias value consistent with the estimate of the signal-to-noise ratio.

30   4.   The method of claim 1, wherein calculating k weighted symbol metrics, for each M-ary symbol, further comprises the steps of:

      4A)   selecting k maximum one outputs, one for each of the k
35   data symbols, wherein a maximum one output is a largest of

24

the M outputs for which the corresponding data symbol is a one;

5   4B)   selecting k maximum zero outputs, one for each of the k data symbols, wherein a maximum zero output is a largest of the M outputs for which the corresponding data symbol is a zero;

10   4C)   determining k differences between the k maximum zero outputs and the k maximum one outputs; and

           4D)   multiplying a function of the signal-to-noise ratio by each of the k differences to produce k weighted symbol metrics.

15

           5.   The method of claim 4, wherein at least one of 5A-5C:
                   5A)   in step 4D), the function of the signal-to-noise ratio is an estimate of the signal-to-noise ratio;
                   5B)   in step 4D), the function of the signal-to-noise
20   ratio is a non-linear function of an estimate of the signal-to-noise ratio; and
                   5C)   in step 4D), the function of the signal-to-noise ratio is the square root of an estimate of the signal-to-noise ratio.

25

           6.   The method of claim 1, wherein in step 1C), a Viterbi decoder is used to decode the value for the non-coherently demodulated signal.

30

25

7.   A decoder circuit in a communication system for
efficiently decoding a non-coherently demodulated signal,
wherein the non-coherently demodulated signal includes a
plurality of M-ary symbols each with M outputs and
5   corresponding to k data symbols, where M equals 2^k, and the
non-coherently demodulated signal is associated with a
signal-to-noise ratio, comprising:

7A)   a symbol metric calculator for receiving the signal-to-
10   noise ratio and the M outputs representing symbols, and
calculating k weighted symbol metrics for each M-ary symbol
to produce a plurality of weighted symbol metrics;

7B)   a symbol metric deinterleaver, operably coupled to the
15   symbol metric calculator, for deinterleaving the plurality of
weighted symbol metrics and producing a plurality of
deinterleaved weighted symbol metrics; and

7C)   a decoder, operably coupled to the symbol metric
20   deinterleaver, for receiving the plurality of deinterleaved
weighted symbol metrics and producing a decoded bit sequence
based on each deinterleaved weighted symbol metric.

26

8.   The decoder circuit of claim 7, wherein at least one of 8A-8B:

8A)   in the symbol metric calculator, the signal-to-noise ratio is an estimate; and

5   8B)   the decoder is a Viterbi decoder.

27

9.   A system having a decoder circuit, for efficiently decoding a received signal, wherein the received signal is binary encoded, bit interleaved, and modulated using M-ary modulation, where M equals 2^k and k is a number of data
5   symbols, the receiver system comprising:

9A)   a M-ary demodulator device for producing a non-coherently demodulated signal in response to the received signal;
10

9B)   a square-law combiner, operably coupled to the M-ary demodulator, for producing M outputs from the non-coherently demodulated signal;

15   9C) a signal-to-noise ratio calculator, operably coupled to the square-law combiner, for computing a signal-to-noise ratio in response to the M outputs;

9D)   a delay element, operably coupled to the square-law
20   combiner, for producing M delayed outputs based upon the M outputs representing symbols from the square-law combiner;

9E)   a symbol metric calculator, operably coupled to the signal-to-noise calculator and the delay element, for receiving
25   the signal-to-noise ratio and the M delayed outputs and producing a plurality of weighted symbol metrics;

9F)   a symbol metric deinterleaver, operably coupled to the symbol metric calculator, for producing a plurality of
30   deinterleaved weighted symbol metrics utilizing the plurality of weighted symbol metrics; and

9G)   a decoder, operably coupled to the symbol metric deinterleaver, for producing a decoded bit sequence based on
35   the plurality of deinterleaved weighted symbol metrics.

28

10.   The system of claim 9, wherein at least one of 10A-10B:

10A)   the M-ary demodulator device is comprised of a plurality of M-ary demodulators for demodulating a plurality of multipath components of the received signal; and

5   10B)   the signal-to-noise ratio calculator produces the signal-to-noise ratio based on a predetermined power control group of the received signal.

FIG.1
-PRIOR ART-
100

FIG.2

```
┌─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─┐
│    CALCULATING AN ESTIMATE OF THE SIGNAL  │
│    TO NOISE RATIO BASED ON THE M OUTPUTS  ├─208
└─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─ ─┘
```

                                                            <u>200</u>

┌──┐
│   CALCULATING k WEIGHTED SYMBOL METRICS, │
│   FOR EACH M-ARY SYMBOL, BASED ON THE   ├─202
│   SIGNAL TO NOISE RATIO AND THE M OUTPUTS TO │
│   PRODUCE A PLURALITY OF WEIGHTED SYMBOL METRICS │
└──┘

┌──┐
│   DEINTERLEAVING USING A SYMBOL METRIC   │
│   DEINTERLEAVER, THE PLURALITY OF WEIGHTED │
│   SYMBOL METRICS TO PRODUCE A PLURALITY OF ├─204
│   DEINTERLEAVED WEIGHTED SYMBOL METRICS   │
└──┘

┌──┐
│   DECODING, USING A DECODER, THE PLURALITY │
│   OF DEINTERLEAVED WEIGHTED SYMBOL METRICS ├─206
│   TO PRODUCE A DECODED BIT SEQUENCE   │
└──┘

FIG.3

                                                            <u>300</u>

┌──┐
│   SELECTING k MAXIMUM ONE OUTPUTS,   │
│   ONE FOR EACH OF THE k DATA SYMBOLS   ├─302
└──┘

┌──┐
│   SELECTING k MAXIMUM ZERO OUTPUTS,   │
│   ONE FOR EACH OF THE k DATA SYMBOLS   ├─304
└──┘

┌──┐
│   DETERMINING k DIFFERENCES BETWEEN THE k MAXIMUM │
│   ZERO OUTPUTS AND THE k MAXIMUM ONE OUTPUTS ├─306
└──┘

┌──┐
│   MULTIPLYING A FUNCTION OF THE SIGNAL- │
│   TO-NOISE RATIO BY EACH OF THE k   ├─308
│   DIFFERENCES TO PRODUCE k WEIGHTED SYMBOL METRICS │
└──┘

FIG.4

402

OUTPUTS
FROM A
NON-COHERENTLY
DEMODULATED
SIGNAL

406

SYMBOL
METRIC
CALCULATOR

408

410

SYMBOL
METRIC
DEINTERLEAVER

412

414

400

DECODER

DECODED
BIT
SEQUENCE

416

404

SNR

FIG.5

504

506

502

502

500

516

BASE STATION

510

RECEIVER

DECODING
CIRCUIT

400

512

TRANSMITTER

508

PUBLIC
TELEPHONE
NETWORK

514

DECODED BIT SEQUENCE

RECEIVED SIGNAL

FIG. 6

600

5/5

FIG.7

DEFINING AN ESTIMATE OF SIGNAL PLUS
NOISE AS A MAXIMUM OF THE M OUTPUTS ⌐702

700

DEFINING AN ESTIMATE OF THE NOISE
AS THE AVERAGE OF ALL OF THE M
OUTPUTS EXCEPT FOR THE MAXIMUM OF THE M OUTPUTS ⌐704

CALCULATING AN ESTIMATE OF THE SIGNAL
AS THE ESTIMATE OF SIGNAL PLUS
NOISE MINUS THE ESTIMATE OF THE NOISE ⌐706

CALCULATING AN ESTIMATE OF THE SIGNAL-
TO-NOISE RATIO BY DIVIDING THE ESTIMATE
OF THE SIGNAL BY THE ESTIMATE OF THE NOISE ⌐708

CALCULATING AN APPROXIMATELY UNBIASED
SIGNAL-TO-NOISE RATIO ESTIMATE BY
SUBTRACTING, FROM THE ESTIMATE OF THE
SIGNAL-TO-NOISE RATIO, A BIAS
VALUE CONSISTENT WITH THE ESTIMATE
OF THE SIGNAL-TO-NOISE RATIO ⌐710

A.	CLASSIFICATION OF SUBJECT MATTER

IPC(6)   :H04B 1/69
US CL   : 375/200

According to International Patent Classification (IPC) or to both national classification and IPC

B.	FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

   U.S. :   375/200,201,202,203,204,205,206,207,208,209,210; 380/34; 370/18

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C.	DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US,A,4,933,952 (ALBRIEUX ET AL) 12 JUNE 1990.	1-10
A	US,A,5,001,723 (KERR) 19 MARCH 1991.	1-10

☐   Further documents are listed in the continuation of Box C.   ☐   See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be of particular relevance		
"E"	earlier document published on or after the international filing date	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
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"O"	document referring to an oral disclosure, use, exhibition or other means		
"P"	document published prior to the international filing date but later than the priority date claimed	"&"	document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
07 SEPTEMBER 1995	25 SEP 1995

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